ADAPTIVE CHANNEL ESTIMATION IN DS-CDMA DOWNLINK SYSTEMS

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ABSTRACT

We design an adaptive channel estimator (ACE) in the DS-CDMA downlink that has a common pilot channel (CPICH). The proposed ACE first estimates the channel condition including the maximum Doppler frequency without requiring exact a priori information on the operating condition. Then, it adjusts the impulse response of the channel estimation filter in response to the estimated channel condition. The performance of the proposed ACE is verified by computer simulation. The performance improvement with the use of the proposed ACE increases as the channel experiences severe frequency selective fading or the mobile is getting closer to the cell boundary.

I. INTRODUCTION

A predetermined symbol is generally transmitted through a common channel, called common pilot channel (CPICH), for the purpose of synchronization and channel estimation in the downlink of the DS-CDMA system [1,2]. Since the signal power of the CPICH is relatively high, the use of a simple fixed channel estimator (FCE) can provide good channel estimation performance under moderate fading channel condition. However, when the channel experiences severe frequency selective fading or the mobile is getting closer to the cell boundary, the performance of the FCE may be degraded noticeably. This performance degradation can be alleviated with the use of an adaptive channel estimator (ACE). Since the channel condition is time-varying and location-dependent, the cut-off frequency of the channel estimation filter (CEF) needs to be adjusted in real-time in response to the channel variation.

There have been proposed a number of adaptive schemes for channel estimation, including the least mean square (LMS) adaptation and recursive least square (RLS) adaptation method [3]. When the signal to interference power ratio (SIR) is low, these methods may not be applicable since the reference signal used for calculation of the error metric is severely corrupted by the noise. In particular, the RLS scheme requires large implementation complexity. Another approach considers the use of multiple CEFs to choose a CEF according to the estimated speed zone [4,5]. Since these schemes were proposed assuming only Rayleigh fading condition with the classic spectrum and a priori knowledge on the SIR of the received signal, they may have limitation in real applications.

In this paper, we design an ACE by modifying the ACE proposed for the uplink application [6]. The parameters of the proposed ACE are adjusted by estimating the channel condition parameters without a priori information on the operating condition.

Section II describes the channel estimator structure in the DS-CDMA downlink. Section III evaluates the performance of channel estimation with and without the use of the proposed ACE. Finally, conclusions are summarized in Section IV.

II. CHANNEL ESTIMATION IN THE DOWNLINK

Fig. 1 (a) depicts the structure of the transmitter in a DS-CDMA downlink system. The information bit in the dedicated physical channel (DPCH) of the i-th user is spread using a channelization code $c_{d,i}$. The predetermined pilot bit of the CPICH is also spread using a channelization code c_p. Each spread sequence is multiplied by the transmit gain $g_{d,i}$ and aggregated before being scrambled. The gain of the DPCH is adjusted according to the channel condition, while that of the CPICH is fixed. Finally, the aggregated signal is scrambled by a scrambling code s_n and transmitted after pulse-shaping. Although the dedicated pilot symbol in the DPCH may also be used for channel estimation as well as power control, it is not used for channel estimation in this paper, since the transmit power of the DPCH is usually much smaller than that of the CPICH.

Assuming that the large-scale fading is perfectly compensated by slow power control, we consider only the small-scale fading. Assuming also that the signals from each path are independent with each other and their relative delays are spaced by an integer multiple of the chip time, the channel can be modeled as an FIR filter with time-varying coefficients. The impulse response of the channel at time t can be represented as

$$h(t,\tau)=\sum_{l=0}^{L-1} h_l(t)\delta(t-d_l T_c) \tag{1}$$

where T_c is the chip duration, L is the number of the propagation paths, $\delta(t)$ is Dirac delta function, d_l is a nonnegative integer representing the delay of the l-th path and $h_l(t)$ is the impulse response of the l-th path at time t represented by

$$\begin{aligned} h_l(t) &\equiv \alpha_l(t)e^{j\phi_l(t)} \\ &= \alpha_{d,l}(t)e^{j\phi_{d,l}(t)}+\alpha_{r,l}(t)e^{j\phi_{r,l}(t)}. \end{aligned} \tag{2}$$

Here, $\alpha_l(t)$ and $\phi_l(t)$ are the amplitude and the phase response of the l-th path at time t, respectively, $\alpha_{d,l}(t)$ and $\phi_{d,l}(t)$ respectively denote the amplitude and the phase response of the direct-path of the l-th path, and $\alpha_{r,l}(t)$ and $\phi_{r,l}(t)$ are those of the scattered components. We assume that $\alpha_{r,l}(t)$ is Rayleigh distributed with average power σ_l^2 and $\phi_{r,l}(t)$ is uniformly distributed over $[0 \;\; 2\pi)$.

Letting P_l and θ_l be the power and the arrival angle of the direct-path ray of the l-th path signal, respectively, the direct-path component can be represented as

$$\alpha_{d,l}(t) = \sqrt{P_l}$$
$$\phi_{d,l}(t) = 2\pi f_d t \cos(\theta_l) \tag{3}$$

where f_d denotes the maximum Doppler frequency. Assuming that θ_l and P_l are unchanged within a few hundred milliseconds interval, it can be modeled that the amplitude $\alpha_l(t)$ has a Ricean distribution with Ricean factor K_l equal to P_l/σ_l^2 and the phase $\phi_l(t)$ has a uniform distribution over $[0 \;\; 2\pi)$. We also assume that the channel gain is normalized, *i.e.*, $\sum_{l=0}^{L-1} E\{|h_l(t)|^2\} = 1$.

Fig. 1 (b) depicts the l-th finger of the rake receiver in the mobile for demodulation of the DPCH signal, where $r_l[kS_{f,d}+n]$ is the received signal $r(t)$ sampled at the $t = kT+nTc$ in the l-th finger. Here, $S_{f,d}$ is the spreading factor of the DPCH and $r(t)$ can be represented as

$$r(t) = \sum_{l=0}^{L-1} h_l(t)x(t - d_l T_c) + \xi(t) + n(t) \tag{4}$$

where $x(t)$ is the transmit signal, $\xi(t)$ and $n(t)$ denote the received intercell interference and the background noise term, respectively. Let *Ior* and *Ioc* be the average received power of the intracell signal (*i.e.*, all the signals from the desired base station) and the intercell interference, respectively, *i.e.*, $E\{|x(t)|^2\} = Ior$ and $E\{|\xi(t)|^2\} = Ioc$. The intercell interference can be approximated as an additive white Gaussian noise term by the central limit theorem.

The received signal is multiplied with the DPCH channelization code and then accumulated for an amount of $S_{f,d}$ samples to obtain the despread data symbol. The CPICH signal is accumulated for an amount of $S_{f,c}$ samples, where $S_{f,c}$ is the spreading factor of the CPICH. Then, it is multiplied with the predetermined pilot symbol to obtain the despread pilot symbol $\tilde{h}_l[k]$. Since $\tilde{h}_l[k]$ is corrupted by the noise, it needs to reduce the noise effect. The channel estimator provides the channel information $\hat{h}_l[k]$ by filtering $\tilde{h}_l[k]$ using a CEF. The CEF can be implemented in the form of a finite impulse response (FIR) filter or infinite impulse response (IIR) filter. When an N-tap FIR filter is used as the CEF, the channel information can be obtained by

$$\hat{h}_l[k] = \sum_{i=0}^{N-1} c_i \tilde{h}_l[k-i], \quad n_d = (N-1)/2 \tag{5}$$

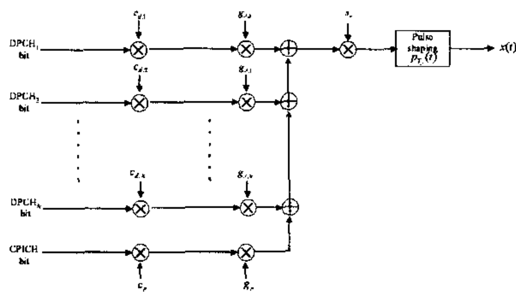
where c_i is the i-th tap coefficient of the FIR filter and n_d is the estimation delay due to the CEF. Note that the moving average (MA) FIR filter is a special case that $c_i = 1/N$ for all i. When an one-pole IIR filter is used as the CEF, the channel information can be obtained

$$\hat{h}_l[k] = \alpha\hat{h}_l[k-1] + (1-\alpha)\tilde{h}_l[k] \tag{6}$$

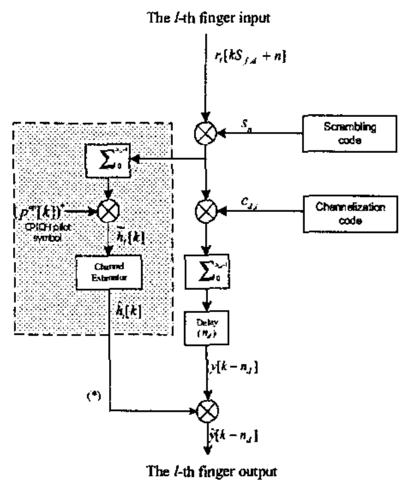
where α is the forgetting factor. Note that n_d in (5) is set to a value of $(1-\alpha)^{-1}$ symbols corresponding to the group delay of the one-pole IIR filter at $f = 0$. The characteristics of a desired CEF can be obtained by controlling either the forgetting factor of the one-pole IIR filter or the tap size and coefficients of the FIR filter.

III. PERFORMANCE EVALUATION

We consider the channel estimation in the WCDMA system as an example [1]. The receiver performance in the WCDMA downlink is evaluated in terms of the BER by computer simulation when an one-pole IIR filter or FIR filter is used as the CEF. The simulation is performed under multipath fading channel conditions with no use of transmit antenna diversity and handover. The simulation condition is summarized in Table 1 and 2 [7], where *Ec* and *Ec,p* denote the chip energy of the received DPCH and CPICH signal, respectively. A convolutional code is used at a data rate of 12.2 Kbps and a turbo code is used at a data rate of 144 Kbps.



(a) The transmitter



(b) The l-th finger of a rake receiver

Fig. 1 A baseband structure of the DS-CDMA downlink system.

A. Conventional channel estimator

Fig. 2 depicts additionally required DPCH *Ec/Ior* as a function of *Ec,p/Ior* to obtain a BER of 10^{-3} at a transmission rate of 12.2 Kbps compared to the use of the ideal channel estimator under the Case-1 and -3 channel conditions with *Ior/Ioc*=-3dB and 9dB without frequency offset. Here, the ideal channel estimation indicates the case when the channel impulse response is perfectly known. We first consider the use of a simple one-pole IIR filter as the CEF. The forgetting factor of the one-pole IIR CEF is set to 0.6, 0.8, 0.9 and 0.95 (corresponding to an averaging time interval of 0.5, 1, 2, and 4 slots, respectively) and f_d is set to 5.5Hz. It can be seen that the additional amount of DPCH *Ec/Ior* decreases as *Ec,p/Ior* increases since the channel information is getting accurate and it increases as *Ior/Ioc* decreases since the despread CPICH symbol is much

Table 1. Simulation condition

	Parameter	Value
DPCH	Data rate	12.2 Kbps, 144 Kbps
CPICH	*Ec,p/Ior*	-16 dB ~ -8.5 dB
Channel	Rayleigh	3GPP model (Case-1, Case-3)
CEF		One-pole IIR, MA FIR, General FIR
Power control		Step size: 1dB, period: 1 slot (0.667ms)

Table 2. Propagation channel conditions

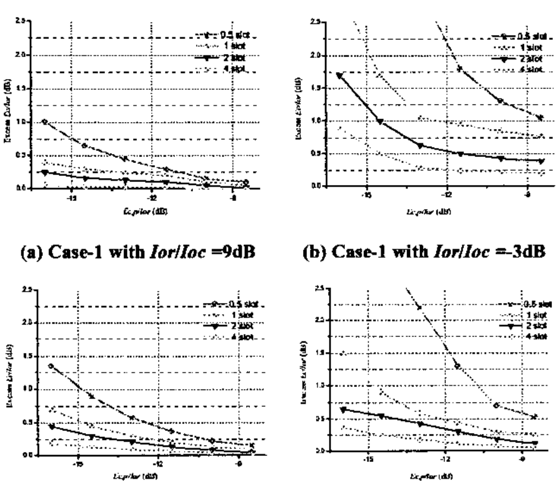
3GPP Case-1		3GPP Case-3	
Delay [ns]	Power [dB]	Delay [ns]	Power [dB]
0	0	0	0
976	-10	260	-3
		521	-6
		781	-9

corrupted by the intercell interference that experiences different fading from the intracell signal.

When *Ior/Ioc* is high, the channel condition becomes similar to that of a single-cell environment in the DS-CDMA downlink system. In this case, the SIR after the combining becomes constant regardless of the fading condition since both the desired CPICH signal and the interfering signal experience the same fading. As a result, it may not be possible to obtain the diversity effect due to the increase of the channel paths. As the number of channel paths increases, the signal power is split into the corresponding number and the inter-path interference increases due to the corruption of the orthogonal property, reducing each path SIR of the CPICH symbol. This implies that an additional DPCH *Ec/Ior* is required in a frequency selective fading channel compared to that in a flat fading channel as seen in Fig. 2 (a) and (c). On the other hand, the intracell signal and othercell signal suffer from different fading. This may cause the SIR of the CPICH symbol to have a large variation in time as *Ior/Ioc* decreases. That is, the SIR can suddenly drop off (*i.e.*, in deep fading), which happens often when the channel has a small number of multipaths. Burst bit errors can occur when both the SIR of the CPICH and the DPCH symbol are dropped off. The SIR of the DPCH symbol can be maintained by fast power controlling the DPCH signal. However, the SIR of the CPICH symbol cannot be adjusted since the transmit power of the CPICH is fixed in the downlink. Thus, the channel information becomes more corrupted by the interference, degrading the BER performance. This degradation becomes larger as the maximum Doppler frequency decreases since the channel coding cannot sufficiently compensate the deep fading effect. Therefore, when *Ior/Ioc* is low, the receiver needs more additional DPCH *Ec/Ior* compared to the ideal channel estimation when the channel has a small number of multipaths as seen in Fig. 2 (b) and (d). Unlike in the uplink where the dedicated pilot signal is power controlled according to the channel variation, the performance degradation due to channel estimation in the downlink significantly varies depending upon *Ior/Ioc* (*i.e.*, the location of the mobile station), particularly when the fading condition is getting closer to the flat fading. In order to reduce the performance degradation, it may be desirable to use a pilot symbol in the DPCH as well as the CPICH, since the transmit power of the DPCH is fast power controlled according to the channel variation. Otherwise, the filtering interval of the CEF should be sufficiently long enough to enhance the SIR of the filtered channel information.

It can also be seen from Fig. 2 that the forgetting factor needs to be optimized according to the channel condition including f_d. When f_d is very low, the performance can be improved by using a large α. As f_d increases, however, the forgetting factor should be reduced. To see the effect on the receiver performance of f_d and the filtering time interval, Fig. 3 plots the amount of additional DPCH *Ec/Ior* compared to the ideal channel estimator for different values of f_d under the Case-3 condition. Among four CEFs, the optimum average time interval is 4, 4, 4, 2, 2 (or 1), 1, and 0.5 slots when f_d is 5.5, 20, 55, 111, 222, 500, and 750 Hz, respectively. We can see that the performance of channel estimation is significantly affected by the bandwidth of the CEF and the channel condition including f_d. Note that the amount of additional DPCH *Ec/Ior* has similar shape regardless of the data rate.

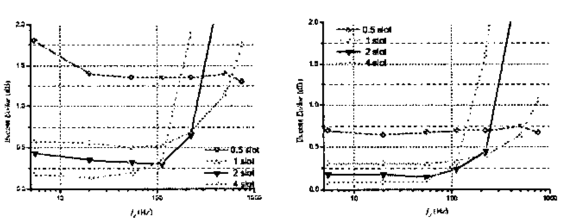
Assume that the system operator provides services to the mobile stations with f_d of up to 750Hz (*i.e.*, 300km/h with 0.1 ppm carrier offset at 2GHz carrier frequency [7]). The CEF with an averaging time interval of one slot (*i.e.*, 0.677 msec) can provide relatively good receiver performance when f_d is between 250Hz



(a) Case-1 with *Ior/Ioc* =9dB **(b) Case-1 with *Ior/Ioc* =-3dB**

(c) Case-3 with *Ior/Ioc* =9dB **(d) Case-3 with *Ior/Ioc* =-3dB**

Fig. 2 Additionally required *Ec/Ior* with 12.2 Kbps and f_d = 5.5Hz.



(a) *Ec,p/Ior*=-13dB **(b) *Ec,p/Ior*=-10dB**

Fig. 3 Additionally required *Ec/Ior* at *Ior/Ioc* = -3dB under the Case-3 condition.

and 600Hz. Thus, we consider the use of a CEF with one slot averaging time interval in the FCE for performance comparison with the use of the proposed ACE. Fig. 4 plots the required DPCH *Ec/Ior* to obtain a BER of 10^{-3} using an one-slot average MA or one-pole IIR filter as the CEF in the Case-3 condition. It can be seen that the performance difference between the use of one-pole IIR CEF and MA FIR CEF becomes larger when f_d is higher than 100Hz since the shape of the frequency response of the two filters becomes different. Therefore, we consider only the use of an FIR filter as the CEF in the proposed ACE.

B. Proposed channel estimator

The simulation results indicate that the performance of channel estimation can be improved if the CEF is properly designed according to the channel condition. When the CPICH is employed as the phase reference, the structure of the channel estimator in the downlink is similar to that in the uplink, where continuous pilot signal is transmitted in parallel with the data signal. This implies that the channel estimation algorithm proposed for the uplink can be applied to the downlink with small modification of the operating parameters [6].

The block diagram of the l-th branch of the proposed ACE is depicted in Fig. 5 [6]. The proposed ACE consists of the channel estimation controller (CEC), the channel parameter estimator (CPE) and the CEF module. The CEC determines the design parameters of the CPE and CEF module considering the operating channel characteristics. Fig. 5 (b) depicts the block diagram of the CPE that estimates the parameters of the channel and transceiver using the received CPICH symbol. The output of the CPE controls the bandwidth of the CEF.

As depicted in Fig. 5 (b), the despread pilot symbol $\tilde{h}_l[k]$ is first lowpass filtered to reduce the noise effect. The filtered output is input to the G_l number of correlators in parallel. The i-th correlator of the l-th branch, i=1,2,..., G_l, correlates the prefiltered pilot symbol with the $m_{l,i}$-sample delayed one for an interval of J symbols. Then, the output of the i-th correlator is normalized as

$$w_l(m_{l,i}) \equiv \frac{\sum_J \mathrm{Re}\{\bar{h}_l^*[k]\bar{h}_l[k-m_{l,i}]\}}{\sum_J |\bar{h}_l[k]|^2} \tag{7}$$

where $m_{l,1} < m_{l,2} < \cdots < m_{l,G_l}$. Using the property that $w_l(m)$ fast decreases as m increases, the channel environment can be classified by comparing $w_l(m)$ with a threshold η. If the j-th correlator output of the l-th finger becomes less than η for the first time, *i.e.*,

$$w_l(m_{l,j}) < \eta \text{ and } w_l(m_{l,i}) \geq \eta,\ i=1,2,\ \ldots,\ j-1, j=1,2,\ldots,G_l \tag{8}$$

the CPE infers that the channel environment belongs to the channel condition-j. In this case, the tap size of the corresponding MA filter is set to $N_{l,j}$, $1 \leq j \leq G_l$. Note that N_{l,G_l+1} corresponds to the case when no correlator output is smaller than the threshold. This happens when the channel response too slowly varies. In this case, it suffices to set the filter tap size to a maximum value since the optimum tap size is not much changed owing to the effect of fast power control [6]. The parameters of the CPE as $\boldsymbol{m}_l$ and N_l are determined by the CEC [6].

Since the proposed ACE is designed considering possible values of the Ricean factor, path power, *Ec,p/Ior* and *Ior/Ioc*, the channel condition can be estimated without exact a priori information on the operating condition. Fig. 6 depicts required DPCH *Ec/Ior* for a BER of 10^{-3} with the use of the proposed ACE at a transmission rate of 144 Kbps. In this case, G_l = 4, $\boldsymbol{m}_l = \{12, 24, 49, 100\}$ and $N_l = \{5, 9, 15, 28, 40\}$. Here, the maximum tap-size of the MA FIR CEF is set to 40 since the use of a CEF with tap size larger than 40 provides negligible performance improvement. It can be seen that the performance with the use of ACE is better than that of the fixed one (with one slot average) under most of the channel conditions, particularly when f_d is low. Note also that the performance improvement at 12.2 Kbps rate is quite similar to that at 144 Kbps rate.

Although a simple MA FIR filter was used as the CEF in [6], a

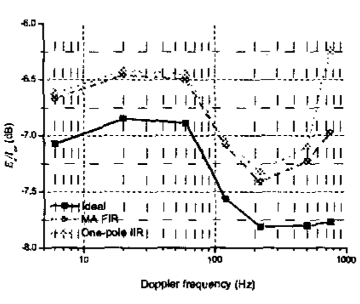
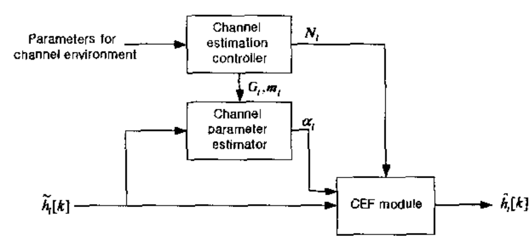
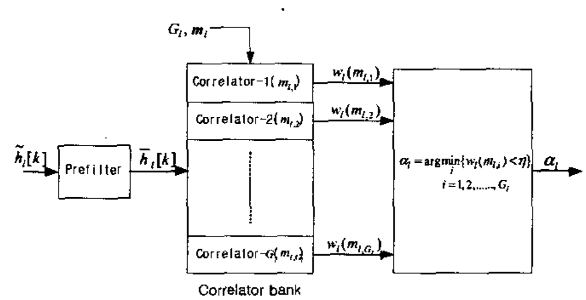


Fig. 4 Required *Ec/Ior* at a BER of 10^{-3} at *Ior/Ioc* = -3dB and *Ec,p/Ior* =-13dB.



(a) The structure of the ACE



(b) The structure of the CPE

Fig. 5 The proposed adaptive channel estimator.

general type FIR filter can be also used in the ACE. The use of a general FIR filter as the CEF requires large implementation complexity, but the performance of the proposed ACE can be improved. The FIR CEF can be optimally designed in the MMSE sense by minimizing

$$\varepsilon_l(v) \equiv E\left\{\left|\hat{h}_l[k]-h_l[k-n_d]\right|^2\right\} \tag{9}$$

where v is the coefficient vector of an N-tap FIR CEF. It is well known that the optimum CEF minimizing (9) is the Wiener filter under stationary channel condition represented by

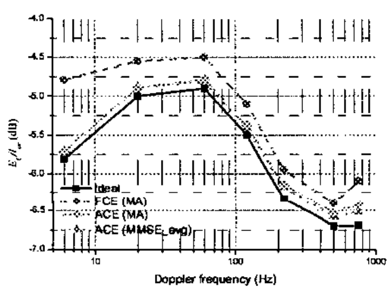
$$v_o = R^{-1}P. \tag{10}$$

Here, $R = E\{X_{l,k}X_{l,k}^H\}$ and $P = E\{h_l[k-n_d]X_{l,k}\}$, where $X_{l,k} = \{\tilde{h}_l[k] \quad \tilde{h}_l[k-1] \quad \ldots \quad \tilde{h}_l[k-(N-1)]\}^T$. However, the Wiener filter may not be realizable in practice since all the parameters on the channel and interference power cannot be known perfectly.

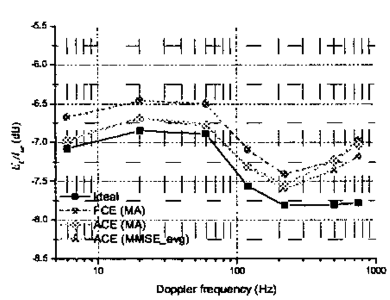
Each MA CEF in the CEF module of the proposed ACE is designed to be optimum under its presumed channel condition [6]. Therefore, we design the FIR CEF based on the Wiener filtering in an average sense. That is, the correlation matrix R and vector P can be obtained by averaging over a possible range of the channel condition,

$$v_a = R_a^{-1}P_a \tag{11}$$

where $R_a = E\{R\}$ and $P_a = E\{P\}$.



(a) Case-1 channel



(b) Case-3 channel

Fig. 6 Required DPCH *Ec/Ior* at a BER of 10^{-3} at *Ior/Ioc* = -3dB and *Ec,p/Ior* = -13dB.

Fig. 6 also depicts the required *Ec/Ior* for a BER of 10^{-3} with the use of 40-tap FIR CEF. It can be seen that the proposed ACE with an averaged MMSE FIR CEF provides a small additional gain over the use of an MA CEF when f_d is larger than 100Hz. This is because as f_d decreases, the frequency response of the FIR CEF becomes similar to that of the MA CEF since the tap size of the corrsponding MA CEF increases. In addition, numerical results show that the receiver performance with the averaged MMSE FIR CEF is slightly poorer than that of the optimum MMSE FIR CEF.

IV. CONCLUSIONS

We have designed an adaptive channel estimator in the DS-CDMA downlink, where the pilot signal in the CPICH is used for channel estimation. The proposed channel estimation scheme is evaluated in terms of the receiver performance under various channel conditions, including the fading statistics, maximum Doppler frequency, *Ior/Ioc*, and data rate. Numerical results show that the performance of channel estimation in the downlink has different behavior from that in the uplink. Since the CPICH signal power is relatively high, a simple fixed channel estimator can provide relatively good channel estimation performance under moderate fading channel condition. However, when the channel experiences severe frequency selective fading or the mobile closes to the cell boundary, the performance of the fixed channel estimator can be degraded significantly. This performance degradation can be alleviated with the use of the proposed ACE. Since the proposed ACE is relatively simple to implement and works without exact a priori channel condition, it can easily be applied to the mobile handset.

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